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Title: .

METHOD AND APPARATUS FOR PRODUCING HIGH-FREQUENCY OSCILLATIONS

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BEAR-P021

METHOD AND APPARATUS FOR PRODUCING HIGH-FREQUENCY OSCILLATIONS

Background Of The Invention

5   Field of the Invention

The invention relates to a method and apparatus for producing high-frequency oscillations. More specifically, the invention relates to a method and apparatus for producing such oscillations without the drawbacks of prior high-frequency oscillators.

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10   Related Art

Almost every electronic instrument contains an oscillator or waveform generator of some type. Such devices as signal generators, function generators and pulse generators clearly require a source of regular oscillations. However, perhaps less obviously, any cyclical measuring instrument or other instrument that initiates

15   measurements or processes or whose function involves periodic states or waveforms also requires such a source. Such devices may include networking equipment, wireless equipment, computers and other applications. In order to allow these devices to operate at higher and higher frequencies, the frequency of the waveform sources must increase as well.

20   It is desirable to use oscillators that maximize noise performance while operating with minimum power consumption, and that cost as little as possible (i.e., take up the least amount of die area). However, at very high frequencies, oscillators may suffer from a number of problems, including:

1.   Resonator losses. Metalization and via losses may lead to increased power

consumption for a given phase noise performance. For example, when

using shorted microstrip transmission lines, a desirable approach at very

high frequencies, such losses associated with the short become particularly

troublesome and degrade the performance of the oscillator.

2.   Layout difficulties. The layout of the oscillator circuit should be very

compact at very high frequencies in order to minimize the influence of

parasitic reactances on oscillator performance. Since the resonator is often

much larger than the active circuitry, such a compact layout may be

difficult to achieve, particularly when the two ports of a resonator are

separated physically.

3.   Noise cyclostationarity. Oscillators typically suffer from periodic

moments of increased or decreased sensitivity to noise from external

sources or from active devices in the oscillator circuit itself. It is highly

desirable to refresh the energy stored in the resonator during such

moments of decreased sensitivity in order to improve the performance of

the oscillator with respect to phase noise and jitter. This approach requires

accurate time-alignment of the active circuitry driving waveforms with

oscillation voltages, which alignment becomes increasingly difficult as the

frequency increases. Such an approach also requires a circuit architecture

that is capable of generating the narrow current pulses that are applied to

the resonator during the moments of reduced sensitivity.

4.   Device delay. The unavoidable delay in active devices makes it difficult

to optimally align the response of the active circuitry with the resonator

waveforms. The resulting phase shift associated with this delay leads to

an off-resonance oscillation and a corresponding reduction of the

5   amplitude of the oscillation, and causes degraded phase noise

performance. Device delay also reduces startup gain margin, possibly

preventing startup entirely, and may also introduce unwanted parasitic

oscillation modes.

In addition, it is often desirable to cross-couple oscillators to generate quadrature

10   signals for timing purposes. One such example is shown in U.S. Patent No. 6,188,292

(see, e.g., Fig. 3). However, this approach uses variable coupling after the non-linear

elements of the oscillators so that the frequency of the output may be varied, and this

results in high phase noise that is undesirable.

These problems are overcome, at least in part, by the present invention.

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Summary of the Invention

The present invention contains three innovations that represent improvements

over the prior art. First is a new resonator architecture that minimizes via losses and

supports a compact layout of active circuitry. The resonator architecture incorporates

20   dual resonant transmission lines to reduce resonator loss and facilitate compact layout.

Second is a method for cross-coupling the oscillations of two oscillators in a way

that compensates for the delay in the active devices of the oscillator, thus permitting

accurate alignment of the active circuitry response with the oscillation waveform. The

cross-coupling of the two oscillators improves phase noise performance and eliminates spurious oscillations.

Third is a specific active circuit architecture that provides very narrow pulses for the operation of the oscillator. This architecture provides for accurate cross-coupling and pulsed-mode operation to improve manufacturing stability and also improves phase noise performance.

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Description of the Drawings

Figure 1 is a diagram showing one embodiment of the resonator architecture of the present invention

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Figure 2 illustrates the effect of device delay in a simplified oscillator loop.

Figure 3 is a diagram showing a pair of oscillators coupled together in one embodiment of the present invention.

Figure 4 is a diagram showing one embodiment of a coupling circuit that may be used with the embodiment of the present invention shown in Figure 3.

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Figure 5 is a diagram showing the cross-coupling between two oscillators of the type shown in Figure 1 in one embodiment of the present invention.

Figure 6 is a diagram of a negative resistance amplifier architecture that may be used in the embodiment of the present invention shown in Figure 1.

Figure 7 is a diagram of a negative resistance amplifier that may be used in the embodiment of the present invention shown in Figure 1.

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Figure 8 is a diagram showing one embodiment of an oscillator according to the present invention.

Detailed Description of the Preferred Embodiment

Referring now to Figure 1, a first embodiment of the resonator architecture of the present invention is shown.

5   Resonator 100 contains two center-tapped transmission lines 102 and 104. These lines serve as resonant elements, with external tuning by variable capacitors 106, 108, 110 and 112. To keep modeling simple, and to make the layout as compact as possible, it is desirable to use straight transmission line segments. To achieve the effect of a shorted line without the associated via losses at the shorted end, a line is used that is

10   twice the length of the shorted line that would normally be used, with the center taps ideally each located at a voltage null at the midpoint of the respective lines.

For example, when the resonator is in operation, nodes A and B on transmission line 102 in Figure 1 operate out of phase so that node X, the point of symmetry in the middle of the line, sits at a voltage null, i.e. a "virtual" ground. As a result, connections

15   to node X nominally conduct no signal current, and thus losses normally associated with such connections (such as via losses) do not impair the operation of the resonant line. Though conducting no signal currents, the center tap connection does provide a path for DC current to flow to the power supply (V_{cc}) from the active circuitry located at both ends of the transmission line. The frequency of the resonator may be changed by simply

20   changing the length of the transmission lines.

The use of a straight transmission line causes the two ends of the line, nodes A and B, to be widely separated, which complicates layout of the active circuitry. To overcome this problem, a second transmission line 104 is introduced, running between

nodes C and D, and is operated in such a way that nodes A and D are in phase with each other, and out of phase with respect to nodes B and C. Thus, transmission line 104 is operated in opposite phase to transmission line 102.

With the second transmission line 104 in operation, two sets of active circuitry
5   114 and 116 can be used, one at each end of the layout. Because transmission lines 102 and 104 can be placed very close to each other, the layout of the active circuits 114 and 116 may also be made very compact, thus minimizing parasitic reactances that would otherwise compromise high frequency operation.

When using this structure, it is important to eliminate undesirable oscillation
10   modes that may be supported. Any oscillation mode may be decomposed into a superposition of four distinct modes, which are indicated in Table 1 below:

TABLE 1

	Node A	Node B	Node C	Node D
Parasitic Mode M_{oo}	↑	↓	↓	↑
Parasitic Mode M_{eo}	↑	↑	↓	↓
Parasitic Mode M_{oe}	↑	↓	↑	↓
Parasitic Mode M_{ee}	↑	↑	↑	↑

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20   Since all other nodes can be compared to node A, we can say that node A will always be considered to be "in phase." From this definition, in Table 1 an upward arrow indicates operation in phase with node A, and a downward arrow indicates operation out of phase with node A (thus all arrows for node A are upward).

The desired mode is the primary mode described above, in which nodes A and D are in phase, and nodes B and C are out of phase. For convenience, we denote this mode, the first line of the table, as mode M_{oo}, representing the mode as "odd-odd." The first "odd" means that signals at opposite ends of transmission lines 102 (nodes A-B) and 104

5   (nodes C-D) are out-of-phase with each other, and the second means that the active circuitry blocks 114 and 116 operate in "odd" or "differential" mode such that the terminal signals of the active circuitry are out-of-phase with each other.

The M_{oo} mode is the preferred oscillation mode because resistive elements in the resonator that couple to the resonator at points of symmetry (nodes X, Y, P and Q) do not

10   conduct signal currents in this mode. Such elements include resistors at the center taps of the lines, and resistors at the center tap points of the tuning capacitors at either end of the lines. The remaining resistors couple to the top plates of the tuning capacitors to provide DC bias for those elements. Minimizing losses due to these resistors is accomplished by making their resistances high enough so that they can conduct only a small fraction of the

15   signal current in the resonator.

All of the other modes shown in Table 1 are undesirable. In each of these "parasitic" modes, either the transmission lines 102 and 104, or the circuit blocks 114 and 116, or both operate in "even-mode" fashion where their terminal signals are in-phase. The resonator structure of the present invention eliminates potential oscillation in each of

20   these unwanted modes.

In contrast to the M_{oo} mode, the other modes cause the center-tapped resistive elements to conduct significant portions of the signal currents, and it is the damping action of these elements that eliminates oscillations in the unwanted modes. For

example, in the M_{eo} mode, nodes A and B operate in phase, and nodes C and D operate out of phase. In this case, the resistance on the V_{cc} line conducts signal currents and damps potential oscillations. Similarly, in the M_{oe} mode, nodes A and C operate in phase, and nodes B and D operate out of phase. Thus, resistance on the V_{tp} line conducts

5   signal currents and damps potential oscillations. Finally, in the M_{ee} mode, all four nodes A, B, C and D operate in phase, bringing all of the resistances in the circuit into the signal path to dampen oscillations.

This discussion indicates that only the M_{oo} mode is capable of sustained oscillations with low loss. Careful selection of the resistive elements of the resonator

10   forces this mode to be enabled during startup, and all other modes to be suppressed. The resonator shown in Figure 1 may be made with any acceptable active circuitry elements 114 and 116. These may be the of the type described below, or may be negative resistance cells as known in the prior art. Note that the center taps need not be exactly at the center point of the transmission lines, or at an exact voltage null for the oscillator to

15   function. However, they should be at points where the voltage is substantially null, as any increase in potential at those points will result in some current flow and thus degraded performance.

A second part of the present invention provides a method for compensating for finite active device delay that might otherwise degrade the phase noise performance,

20   reduce the startup gain margin, or introduce oscillation in undesired modes. As such delays are related to the active electronics in an oscillator rather than the type of resonator, the remedy according to the present invention is useful for many types of resonators, not simply the type disclosed in the foregoing paragraphs.

Figure 2 illustrates the effect of device delay in a simplified oscillator loop. Oscillator 120 is shown as a signal loop containing a frequency-selective gain element 122, a delay element 124, and a non-linear element 126. Together, delay element 124 and nonlinear element 126 represent an active device with inherent delay.

5   Sustained oscillations will only occur under conditions that lead to a phase shift which totals zero around the loop. Because the delay element 124 introduces a negative phase shift, the gain element 122 must compensate with a positive phase shift in order for this condition to be satisfied. This causes the oscillation frequency to shift away from the resonance frequency f_o, and toward a lower frequency f_{osc}. This frequency shift away

10   from resonance reduces the net gain around the loop, lowering the oscillation amplitude and degrading the signal to noise ratio of oscillator 120. Thus, there is increased jitter and phase noise in oscillator 120.

          A related effect may be observed in the time domain behavior of oscillator 120. Ideally, the peak of the current pulse (for example, at the output of the nonlinear block as

15   in a Colepitts oscillator) will be aligned with the maximum or minimum voltage of the oscillator. Where this occurs, the noise associated with the current pulse is at a maximum at the point where the oscillation waveform is least sensitive to noise injection. However, the delay in oscillator 120 causes the current waveform I_p produced by the non-linearity to be misaligned with the oscillation voltage X by a time T_d. This misalignment further

20   degrades the phase noise performance of the oscillator.

          If the loop delay is too large, the oscillator may not start at all due to insufficient loop gain away from resonance. The unwanted delay may also make other undesired

oscillation modes possible. For example, an oscillator that nominally operates in a differential mode may exhibit common-mode oscillations.

Compensating for loop delay requires a mechanism or device that provides positive phase shift to cancel the negative phase shift of the delay element. One way to
5   accomplish this is to insert a non-minimum phase all-pass filter into the loop. However, the loading on the loop of such a filter may be prohibitively high, particularly at high frequencies, and the phase shift may not be well controlled over variations in manufacture.

An alternative approach in accordance with the present invention is to shift the
10   phase of the oscillator signal by coupling the signal from a second, identical oscillator after the gain elements but before the non-linear elements. This technique is shown in Figure 3. In this architecture, oscillators 130 and 132 are coupled together through coupling blocks 134 and 136. Under this condition, oscillations of oscillators 130 and 132 occur in quadrature; in particular, oscillator 132 leads oscillator 130. Thus, by
15   adjusting the coupling coefficient M of coupling blocks 134 and 136, the delayed signal from oscillator 130 can be summed with an appropriately weighted signal from oscillator 132 so that the input to non-linear element 138 is advanced in phase and properly aligned with the oscillation waveform X (the waveform from oscillator 130).

For coupling in the opposite direction, because oscillator 130 lags oscillator 132,
20   an inversion is necessary to achieve the same beneficial phase shift for the non-linear element 140 of oscillator 132. By carefully selecting the coupling coefficient M of coupling blocks 134 and 136, the loop delay can be canceled entirely, allowing both

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oscillators to run at resonance with proper time alignment of the current pulses with respect to the oscillation voltages.

The coupling can be performed anywhere in the loop between the output of the gain element and the input of the nonlinear element. However, maximum benefit is obtained if the signals from oscillators 130 and 132 are coupled in a manner that maintains their quadrature phase relationship in the summing process. For example, any path delays between the signals from oscillators 130 and 132 at a particular summation point should ideally be balanced.

It is also important that the signals from oscillators 130 and 132 be coupled after being filtered by the gain elements, but prior to the nonlinear blocks. Assuming that the gain blocks eliminate significant harmonics of the oscillator signals, summation prior to the nonlinear blocks results in a simple phase shift θ where:

$$\theta = \arctan(M) \text{ radians}$$

where M is the coupling coefficient. This would not necessarily be true if the two signals are coupled after the nonlinear blocks. For example, if the nonlinear blocks produce narrow pulses, then coupling the oscillator signals after the nonlinear blocks would broaden the pulse in addition to producing a phase shift. This broadening may degrade the phase noise performance and loop gain, and thus be undesirable.

A significant advantage of this technique is that the required coupling gain M can be obtained with high precision, even at very high frequencies. This is due to the fact that the gain, like other unitless quantities, can be made to track the ratio of electrical elements, rather than the absolute value of those elements. Thus, for example, the coupling gain M might be set by the ratio of two resistors or capacitors, or by the ratio of

the gains of two active devices. Variations in the absolute component values will not

produce significant deviations in the coupling gain M, as long as the ratio of the values is

tightly controlled. This is easily done in practice, particularly in integrated circuit

technologies, but matched components suited to this purpose are also available in discrete

5   technologies.

Consider the coupling circuit 140 shown in Figure 4, in which two voltages, V1

and V2, are applied to a capacitive summing network comprising two capacitors 142 and

144, having capacitances C1 and C2 respectively, and an element 146 with an unknown

admittance Y. The output of the coupling circuit, V_{sum}, is nominally a weighted sum of

10   the two input voltages. The unknown admittance Y may represent, for example, the

admittance of an active device plus parasitic admittances that are significant at high

frequencies, but that may not be well modeled or may exhibit significant variations in

manufacturing.

Since a constant phase shift is desired, it would be ideal for the phase of the

15   output to be independent of the admittance Y, and this is in fact the case. The total output

voltage is given by:

$$\mathrm{Vsum} = V1 \, \frac{C1}{C1 + C2 + Y/s} + V2 \, \frac{C2}{C1 + C2 + Y/s}$$

20   If V1 and V2 are in quadrature, with equal amplitudes, and V2 leading V1, then

the phase shift θ is given by:

$$\theta = \arctan \frac{C2}{C1}$$

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Thus, the phase of the output depends only upon the ratio of the two capacitances, and is independent of the admittance Y.

Figure 5 is a diagram of two oscillators 170 and 172, each of the type shown in Figure 1, coupled together in this fashion. Due to the fact that oscillators 170 and 172

5   each incorporate two negative resistance cells, it is necessary to couple the oscillators together at both ends of the transmission lines.

In addition to a coupling network to couple two oscillators together, it is advantageous to employ a negative resistance cell that produces narrow current pulses in response to the coupled signals to periodically refresh the energy stored in the resonator.

10   As described in the foregoing discussion, these pulses should ideally be aligned with the maxima or minima of the resonator signals to minimize the effect of noise fluctuations associated with the current pulses. An example of a single-ended negative resistance cell that produces pulses with such alignment is known in the art as a Colepitts oscillator. The Colepitts architecture is unsuitable for the present invention due to its single-ended

15   nature. What is required is an alternative architecture that lends itself to differential operation. An architecture for a negative resistance cell achieving these goals is illustrated in Figure 6.

The architecture makes use of two coupling networks, 182 and 184, and two pulse forming networks, 186 and 188, to form a quasi-differential negative resistance cell, 180.

20   A portion of resonator 100 from Figure 1 is shown in Figure 6 for reference, including transmission lines 102 and 104, variable capacitors 110 and 112, and nodes B and D. The coupling networks couple the resonator signals from the two coupled oscillators to the inputs of the pulse forming networks. The pulse forming networks produce narrow

current pulses in response to their inputs, and these pulses appear at the outputs of the pulse forming networks in approximate alignment with the maxima or minima of the resonator signals. The current pulses are coupled back into the resonator to refresh the energy stored there. As a result of this last coupling, the output of pulse forming network

5   186 is connected to one input of the coupling network 184, which couples to the opposite pulse forming network, 188. Similarly, the output of pulse forming network 188 is connected to one input of the coupling network 182, which couples to the first pulse forming network, 186. This completes a positive feedback loop forming the desired negative resistance cell, 180.

10   The capacitive coupling networks shown in Figure 4 may be used for coupling networks 182 and 184, but alternative coupling networks are also possible. By connecting two coupling networks and pulse forming networks back-to-back in this fashion, the negative resistance cell produces alternating pulses in a "ping-pong" fashion in response to the resonator voltage. This "ping-pong" action occurs because the input

15   signals of the two pulse forming networks are out-of-phase in the desired oscillation mode and thus the pulse forming networks produce streams of pulses that are displaced from each other by half an oscillation cycle.

An example of a circuit using this technique appears in Figure 7. This circuit also incorporates the capacitive coupling network of Figure 4. The unknown admittance Y

20   shown in Figure 4 is present implicitly in the input admittance of the pulse forming network in Figure 7. Thus, Figure 7 shows a simplified schematic of a negative resistance amplifier, 150, that can form one set of the active circuitry shown in the oscillator in Figure 1. When used in conjunction with the resonator of Figure 1, one such

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amplifier appears at each end of the resonator. A portion of resonator 100 from Figure 1 is shown in Figure 7 for reference, including transmission lines 102 and 104, variable capacitors 110 and 112, and nodes B and D.

Capacitors 152, 154, 156 and 158 form a capacitive summing network in

5   accordance with the discussion of Figure 4. Assuming that the signals of the two oscillators are in quadrature, the summed signal at the base of the first pair of transistors 160 and 162 exhibits a well-controlled phase shift that depends primarily on the ratio of the capacitor values, and that compensates for the delay through the other transistors which follow. Alternative embodiments that do not employ coupled oscillators and

10 therefore do not require a summation network would omit some or all of these capacitors. Although the preferred embodiment makes use of capacitors in the summation network, other embodiments are also possible that make use of resistors or active devices to perform the summation, as will be evident to one skilled in the art.

The transistor networks respond to a sinusoidal input drive by producing narrow

15 current pulses that are applied to the resonator. Thus, the transistors form the nonlinear element in the oscillator loop. The resonator filters these pulses, producing a sinusoidal voltage in response that is applied to the capacitive summing networks, thus completing the loop.

This approach differs from the prior art in that the two portions of the circuit

20 operate independently in a "ping-pong" fashion, producing narrow current pulses that can be time-aligned with the minima of the resonator voltage waveform. As described above, this pulsed-mode operation results in reduced oscillator phase noise. In amplifier art, this

may be described at class-AB, class-B or class-C operation, depending upon the conduction angle of the devices.

Conventional differential negative resistance amplifiers produce square wave rectified outputs, rather than narrow pulses, and are described in the art as class-A

5   operation with the exception of the Colepitts oscillator, which also has a pulsed mode of operation. However, as mentioned in previous paragraphs, the Colepitts oscillator generally only appears in single-ended form, and not in the differential circuit of the present invention. Differential circuits are also desirable for their immunity to external noise sources and their lack of even-order distortion products, which produce duty-cycle

10   distortion in an oscillator.

It should be noted, that the architecture of Figure 6 and the associated circuit technique of Figure 7 do not require the use of the resonator of Figure 1, and other types of differential resonators may also be used. Furthermore, the architecture of Figure 6 and the associated circuit technique of Figure 7 do not require the use of coupled oscillators

15   as disclosed in the foregoing paragraphs and may therefore be used in single oscillators, coupled oscillators according to the present invention, or with other coupling relationships between oscillators that are not contemplated in this invention.

An example of an oscillator that takes advantage of these concepts is shown in Figure 8. In Figure 8, four transmission lines, two for each oscillator, are shown in the

20   middle of the figure. The active circuits are located at either end of the transmission lines, and the cross-coupling of the two oscillators is accomplished with routing channels at the extreme sides of the layout. As shown in Figure 8, this architecture facilitates very tight layout of the active devices at the very edges of the transmission lines.

In the foregoing specification, the invention has been described with reference to specific embodiments thereof. It will, however, be evident that various modifications and changes can be made thereto without departing from the broader spirit and scope of the invention as set forth in the appended claims. For example, other configurations of active

5   circuitry may be used, or other types of coupling circuits between the two oscillators. The specification and drawings are, accordingly, to be regarded in an illustrative rather than a restrictive sense. Therefore, the scope of the invention should be limited only by the appended claims.